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Interim Report

PREFERRED-CIRCUIT TECHNIQUES FOR VARACTOR-TUNED FILTERS

Solomon Hecht

Gerald Kannischak

Airborne Instrument Lab

TECHNICAL REPORT NO. RADC-TR-66-708

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FOREWORD

This report was prepared by the Special Systems and Techniques Department of Airborne Instruments Laboratory (AIL) under Contract AF 30(602)-3583, Project 4540, Task No. 454002, for Rome Air Development Center. The AIL report number is 5591-TDR-5.

The work was performed by Mr. Solomon Hecht under the direction of Mr. Stanley Becker, Section Head, and Mr. Gerald Kanischak, Group Leader.

The RADC Project Engineer was Mr. Hollis J. Hewitt, EMCVI-1.

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This report has been reviewed and is approved.

Approved: *Hollis J. Hewitt*
HOLLIS J. HEWITT
Interf Anal & Control Sec
Vulnerability Reduction Branch

Approved: *Samuel D. Zaccari*
SAMUEL D. ZACCARI
Chief, Vulnerability Reduction Br
Communications Division

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ABSTRACT

This report describes an investigation of circuit techniques to minimize spurious reponses in varactor-tuned filters.

Design equations for the center frequency, bandwidth and insertion loss of single and multiple-resonator filters are presented. Series-resonant filters with one and two resonators were constructed and measurements of their performance characteristics were made. Measurements of the spurious response characteristics of the single-resonator filter were also made.

One- and two-resonator filters were also constructed using a pair of opposingly biased varactors (push-pull circuit) for each resonator. Measurements show that such a circuit can significantly reduce the second-harmonic generation and the response to a signal at one half the tuned frequency.

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I. INTRODUCTION

This report describes the work performed to study interference effects in varactor-tuned filters for receiver applications. The study also provides circuit techniques which are useful in reducing interference susceptibility to interfering signals.

The design of electronically tunable filters is usually accomplished by using a varactor or a YIG as the tuning element. The YIG however, is only practical at microwave frequencies, and consequently the varactor is used in electronic tuning applications up to UHF. Associated with varactor-tuned filters are interference properties resulting from the nonlinear characteristics of the varactor. In particular, properties such as harmonic generation, intermodulation, cross modulation, detuning and oscillations arise. When a filter of this type is used as an RF preselector for a receiver, these interference properties can seriously degrade the receiver performance. Therefore, it is important to eliminate or to at least reduce these unwanted properties, in order to make the varactor-tuned filter a useful receiver component.

This report evaluates the series- and parallel-resonant circuit, and shows the advantage of the series-resonant circuit. Design equations are presented for the center frequency, bandwidth, and insertion loss of the single-resonator filter. These equations are then generalized for the multiresonator filter.

Measurements were made to determine the intermodulation and cross-modulation properties of the varactor tuned filter. A varactor filter was then constructed using a pair of opposingly-biased varactors (analogous to a push-pull circuit). Measurements show that the latter circuit had reduced susceptibility to even-order intermodulation.

II. DESIGN

A. PROPERTIES OF SERIES AND PARALLEL RESONANT CIRCUITS

We now consider the two basic resonant circuits, the series and the parallel RLC circuits. Either circuit can be used as a filter element and we must therefore, choose whichever configuration is most applicable for our design. The properties that are of interest are variation of bandwidth and insertion loss as a function of center frequency. Generally, constant bandwidth and constant insertion loss over the tuning range are desirable.

1. PARALLEL RLC TUNED CIRCUIT

The varactor-tuned parallel RLC circuit is shown in Figure 1A.
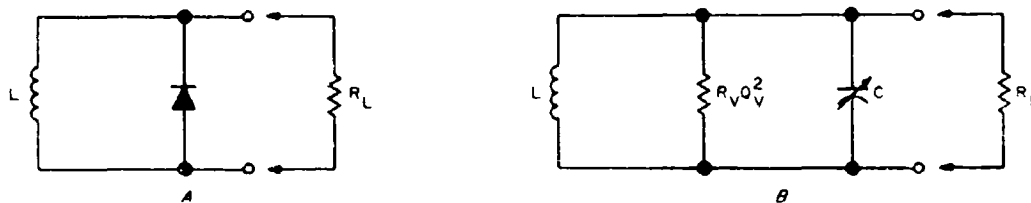
FIGURE 1.   PARALLEL-TUNED RESONANT CIRCUIT

Using the equivalent circuit for the varactor, the circuit in Figure 1A can be approximated by the circuit shown in Figure 1B where

R_V = varactor series resistance

Q_V = varactor quality factor $\omega R_V C$

The center frequency of this circuit is given by:

$$f_o = \frac{1}{2\pi \sqrt{LC}}$$

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With C_{max} and C_{min}, the maximum and minimum capacitance values respectively, the center frequency can be varied from f_{min} to f_{max} where

$$f_{min} = \frac{1}{2\pi \sqrt{LC_{max}}} \; ; \; f_{max} = \frac{1}{2\pi \sqrt{LC_{min}}}$$

The tuning range is the ratio of the maximum to minimum frequency. Thus,

$$\frac{f_{max}}{f_{min}} = \left[\frac{C_{max}}{C_{min}} \right]^{\frac{1}{2}}$$

The 3-db bandwidth of a resonant circuit is given by

$$f_{3db} = \frac{f_o}{Q_L}$$

where Q_L is the circuit loaded-Q given by

$$Q_L = \frac{R_v Q_v^2 \mid\mid R_L}{\omega_o L}$$

When high-Q varactors and low impedance loads are used, $R_v Q_v^2 \gg R_L$, and the expression for the loaded Q reduces to

$$Q_L = \frac{R_L}{\omega_o L}$$

The bandwidth therefore, is given by

$$f_{3\text{ db}} = \frac{2\pi f_o^2 L}{R_L}$$

and is seen to be proportional to the square of the center frequency.

The insertion loss for a single resonator is given by

$$\text{I. L. in db} = 20 \log \frac{Q_u}{Q_u - Q_L} = 20 \log \frac{1}{1 - \dfrac{Q_L}{Q_u}}$$

Since

$$Q_L = \frac{R_L}{\omega_o L}$$

and the unloaded circuit Q, Q_u, is given by

$$Q_u = \frac{R_v Q_v^2}{\omega_o L}$$

then,

$$\frac{Q_L}{Q_u} = \frac{R_L}{R_v Q_v^2} = \frac{R_L}{R_v^3 \omega_o^2 C^2} = \frac{\omega_o^4 L^2 R_L}{R_v^3 \omega_o^2} = \frac{R_L L^2}{R_v^3} \omega_o^2$$

Thus, the insertion loss as a function of frequency is given by

$$\text{I. L. in db} = 20 \log \left[\frac{1}{1 - \dfrac{R_L L^2}{R_v^3} \omega_0^2} \right]$$

2. SERIES RLC TUNED CIRCUIT

The varactor-tuned series RLC circuit is shown in Figure 2A, and the equivalent circuit is shown in Figure 2B.
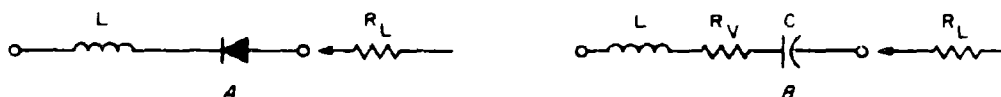
FIGURE 2. SERIES-TUNED RESONANT CIRCUIT

The center frequency and tuning range are the same as for the parallel circuit.

The 3-db bandwidth is given by f_0/Q_L, but Q_L has a different value now. It is given by

$$Q_L = \frac{\omega_0 L}{R_L + R_v}$$

and can be approximated by

$$Q_L = \frac{\omega_0 L}{R_L}$$

for high-Q varactors.

The expression for the 3-db bandwidth, then reduces to

$$f_{3\,db} = \frac{f_0 R_L}{\omega_0 L} = \frac{R_L}{2\pi L} = \text{constant}$$

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Thus, it is seen that the bandwidth does not change with variation in center frequency.

The insertion loss is given by

$$I.\,L. = db = 20\log \cfrac{1}{1 - \cfrac{Q_L}{Q_u}}.$$

$$\frac{Q_L}{Q_u} = \cfrac{\cfrac{\omega_o L}{R_L}}{\cfrac{\omega_o L}{R_v}} = \frac{R_v}{R_L} = \text{constant}$$

Thus, it is seen that the insertion loss remains constant as a function of center frequency and equal to

$$I.\,L. = db = 20\log \cfrac{1}{1 - \cfrac{R_v}{R_L}}$$

Thus, for varactor-tuned filters the series-tuned RLC circuit is ideal. It exhibits constant bandwidth and constant insertion loss over the tuning range.

B.   DESIGN OF SINGLE-RESONATOR VARACTOR-TUNED FILTERS

The equations for a single-resonator varactor-tuned filter will now be derived for a series-tuned circuit with transformer (or magnetic) coupling to the source and load. Capacitive coupling to the source and load is not recommended because the additional capacitance reduces the tuning range of the filter.

The schematic of the filter is shown in Figure 3A.
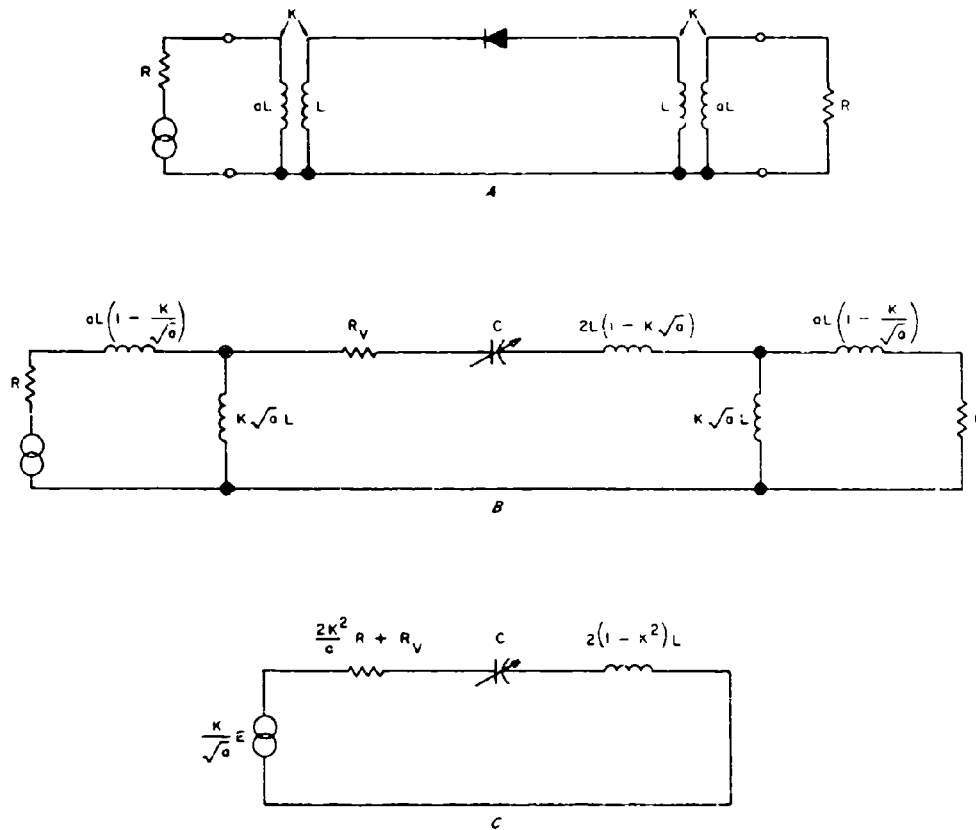
FIGURE 3. SINGLE-RESONATOR TUNABLE FILTER

where

  R = source and load impedances,

  L = secondary inductance,

  a = factor that relates the primary inductance to the secondary inductance,

  K = coefficient of coupling of the transformer.

Using the varactor equivalent circuit and the "T" equivalent for the transformer, the circuit takes the form shown in Figure 3B. By means of Thevenin's theorem, and assuming that $(\omega_0 aL/R)^2 \gg 1$, we can further

reduce the circuit to that shown in Figure 3C. From this circuit, we
see that the center frequency is given by

$$f_o = \frac{1}{2\pi \sqrt{2LC(1 - K^2)}}$$

and the 3-db bandwidth is given by

$$f_{3\ db} = \frac{2RK^2 + a\ R_v}{4\pi aL\ (1 - K^2)}$$

Given a particular varactor and the specifications on the filter, the only
unknowns remaining to be solved are a, K, and L. There are many sets
of values for a, K, and L that will result in the desired filter response,
and therefore, a third equation is necessary to establish the design. One
logical solution would be to design for a particular insertion loss using
the equation:

$$I.\ L.\ in\ db = 20 \log \frac{2RK^2 + aR_v}{2RK^2 + aK^2 R_v}$$

Another solution would be to minimize the number of components required
to construct the filter when the equivalent T's are used for the trans-
formers. (Using equivalent T's to replace transformers in the filter
circuit is advantageous because it eliminates the required control of
mutual coupling between the primary and secondary of the transformer.)
Looking at the relationship of Figure 4, we see that by making $K^2 = 1/a$,
we can eliminate one of the inductors from the equivalent T transformer.

In the design of our filter we used minimum-component
equivalent T transformers to simplify the design.
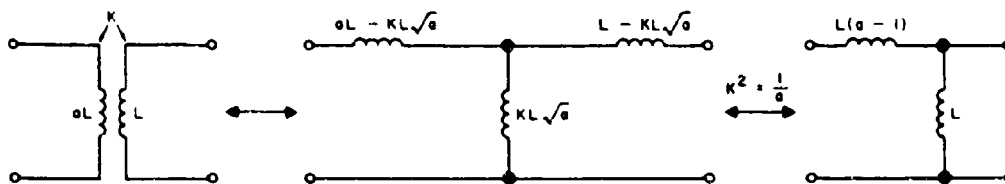
FIGURE 4. TRANSFORMER EQUIVALENT CIRCUITS

C. A DESIGN OF MULTIRESONATOR VARACTOR-TUNED FILTERS

When multiresonator filters have to be designed a particularly useful approach is to use the equal-element filter design (reference 1). The reason for this is that only two voltages are required to tune the filter. One voltage would vary the center frequency of all but the first and last resonator and the other would vary center frequency of the first and last resonator. A Butterworth or Tchebycheff design would require a voltage for every pair of resonators.

The circuit for an arbitrary number of resonators is shown in Figure 5. Since in an equal-element design the filter response is completely specified by (1) the decrements, (2) the inter-resonator coupling and (3) the unloaded Q of the resonators, we shall determine these parameters and obtain design equations. The equivalent circuit of the circuit in Figure 5A with the condition that $K_0^2 = 1/a$ is shown in Figure 5B. The decrement, d, is found by detuning the second resonator and finding the reciprocal of the Q of the equivalent input network. It can be shown that

$$d = \frac{1}{Q_L} = \frac{R_V + RK_0^4}{\omega_0 L (2 - K_0^2)}$$

It can also be shown that the end resonators have a center frequency given by
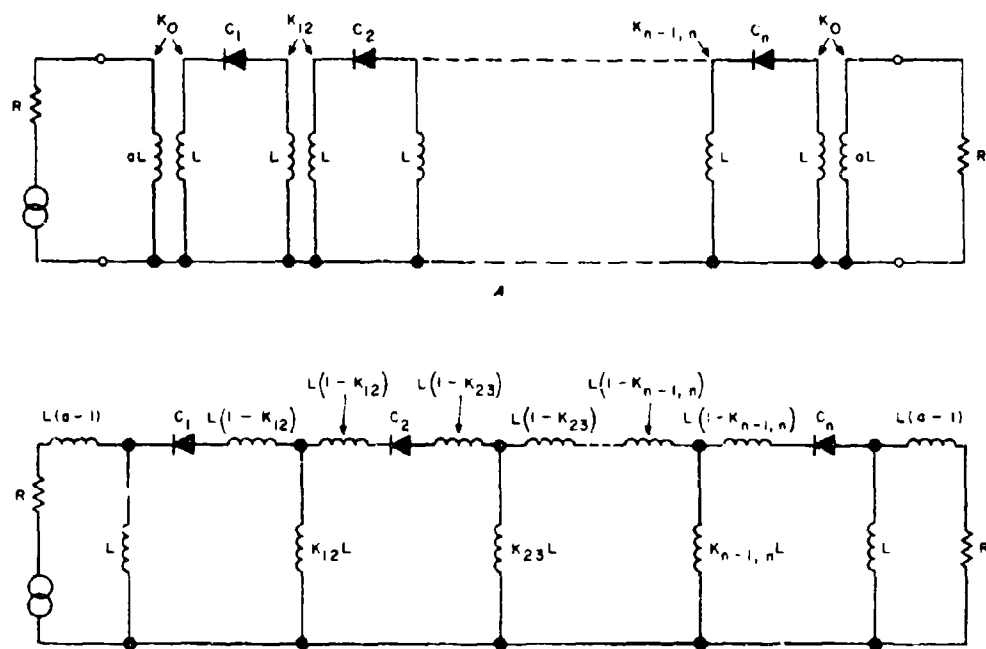
FIGURE 5. MULTIRESONATOR TUNABLE FILTER

$$f_O = \frac{1}{2\pi \sqrt{LC_P (2 - K_0^2)}} \quad P = 1, \, n$$

For all the other resonators the center frequencies must be given by

$$f_O = \frac{1}{2\pi \sqrt{LCq}} \quad q = 2, \, 3, \, - \, - \, - \, -, \, n - 1$$

Thus, we see that at any tuned frequency of the filter, the tuning capacitors of the end resonator are related to the capacitors of the middle resonator by the relationship

$$\frac{C_q}{C_p} = (2 - K_0^2) = r$$

The inter-resonator coupling coefficients K_{12} through $K_{n-1, n}$ are all related to K, which by definition is given by

$$K = d - d_0$$

It can be shown (reference 1) that

$$K_{12} = K_{23} = ----- = K_{n-1, n} = 2K = 2(d - d_0)$$

We now have four definitive equations

$$d = \frac{R_v + RK_0^4}{\omega_0 L (2 - K_0^2)}$$

$$f_0 = \frac{1}{2\pi \sqrt{LCp (2 - K_0^2)}}$$

$$\frac{C_q}{C_p} = (2 - K_0^2) = R$$

$$K_{12} = K_{23} = ----- = K_{n-1, n} = 2(d - d_0)$$

Summarizing, for a particular varactor with capacitance C_p, and given filter specifications of center frequency, bandwidth and tuning range, the unknowns to be solved for are K_0; K_{12}; - - -; $K_{n-1, n}$; L and C_q. With the four definitive equations given above these unknowns can readily be solved.

D.   BIAS CIRCUIT DESIGN

The bias circuit, which must provide the DC tuning voltage for the varactor, is essentially simple. The impedance looking into the bias circuit must be high at all frequencies that the filter must tune to.

Since tuning is accomplished by varying the bias voltage at some finite rate (say a sawtooth) the bias circuit must also be a low impedance for the tuning signal rate. If the highest frequency component of the tuning signal approaches the lowest filter frequency, then a simple RF choke and bypass capacitor will not be adequate and a more elaborate decoupling filter must be used.

Another consideration for the bias circuit is that the bias resistor should be kept small. Large resistors can cause a kind of relaxation oscillation to develop in the bias circuit when the signal voltage is large and the bias voltage is small. Under this condition, the signal becomes modulated at a frequency determined by the bias resistor and the bypass capacitor. Such oscillations have been observed with resistors as small as 10 k when the varactor is biased near zero volt and the signal level is above 0 dbm. If the oscillation is present, increasing the size of the bypass capacitor only reduces the oscillation (or modulation) frequency.

III. EXPERIMENTAL CIRCUITS

A. SINGLE-RESONATOR VARACTOR-TUNED FILTER

A single-resonator varactor-tuned filter was designed using the equations derived in Section II-B. The varactor used in this circuit was a High-Q Varicap made by TRW Semiconductors. The varactor parameters are as follows:

$$C \text{ at } -4 \text{ v} = 14.4 \text{ pf}$$

$$Q \text{ at } 50 \text{ MHz} = 360$$

$$\text{Breakdown voltage} = -87 \text{ volts}$$

The filter was designed for a center frequency of 60 MHz and a 2-MHz 3-db bandwidth when the varactor capacitance is 14.4 pf. The resulting design is shown in Figure 6A. The complete circuit, including the biasing network is shown in Figure 6B. The electrical characteristics of this filter are shown in Figure 7.

B. TWO-RESONATOR VARACTOR-TUNED FILTER

A two-resonator varactor-tuned filter was designed using the equations derived in Section II-C. The varactors used in this circuit were matched Varicaps with the following characteristics:

	Varactor 1	Varactor 2
C at -4 v	15.96 pf	15.96 pf
Q at 50 MHz	> 300	> 300
Breakdown voltage	50 v	50 v

The filter was also designed for a center frequency of 60 MHz and a
2-MHz bandwidth when the varactor capacitance is at 14.4 pf. The
resulting design, including the DC biasing network, is shown in Figure 8. The electrical characteristics of this filter are shown in
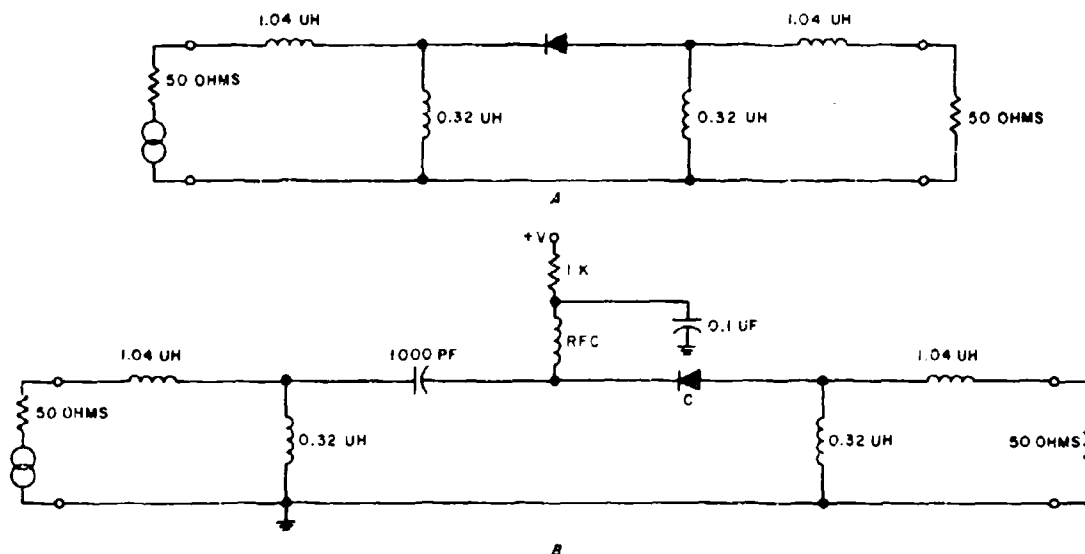Figure 9.

FIGURE 6.   EXPERIMENTAL SINGLE-RESONATOR TUNABLE
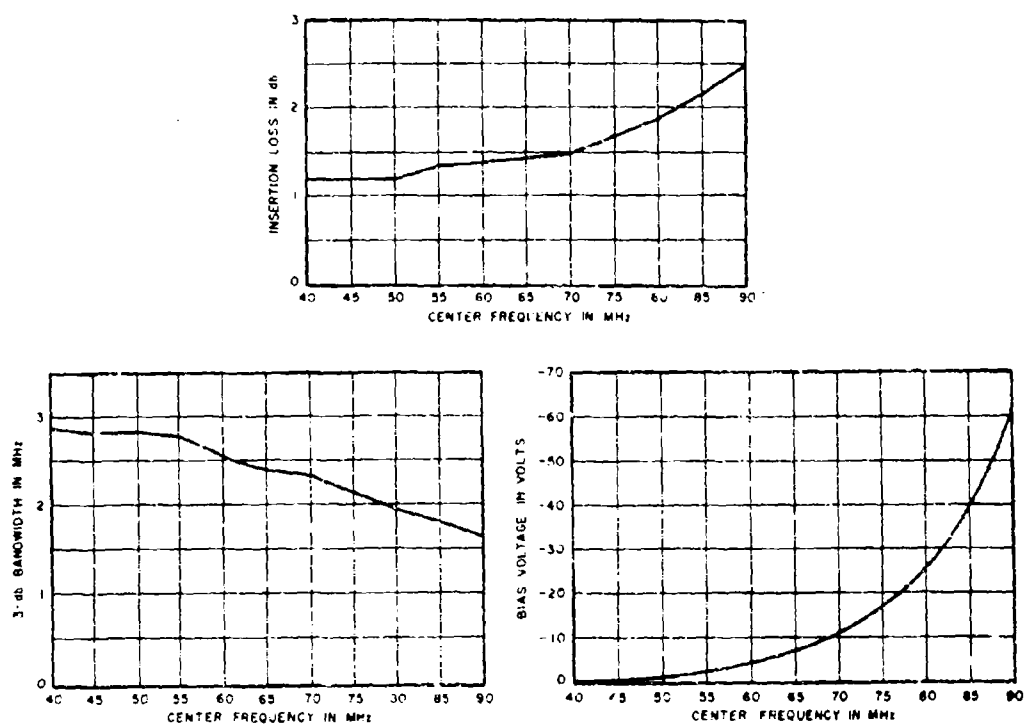   FILTER

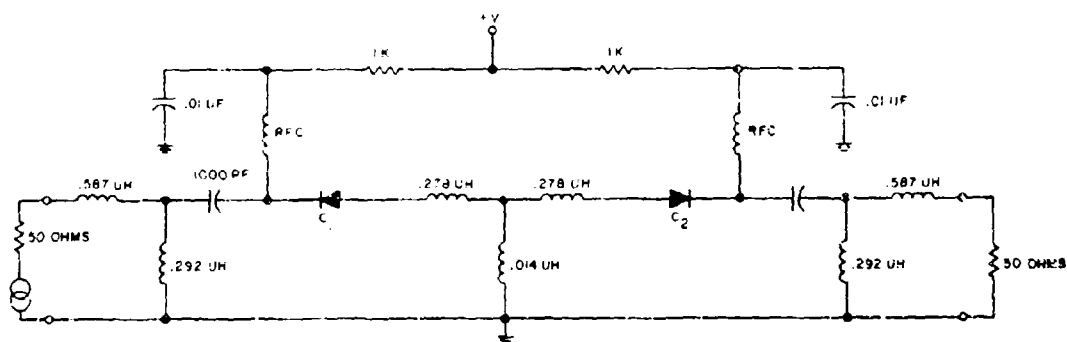FIGURE 7. ELECTRICAL CHARACTERISTICS OF SINGLE-
RESONATOR TUNABLE FILTER

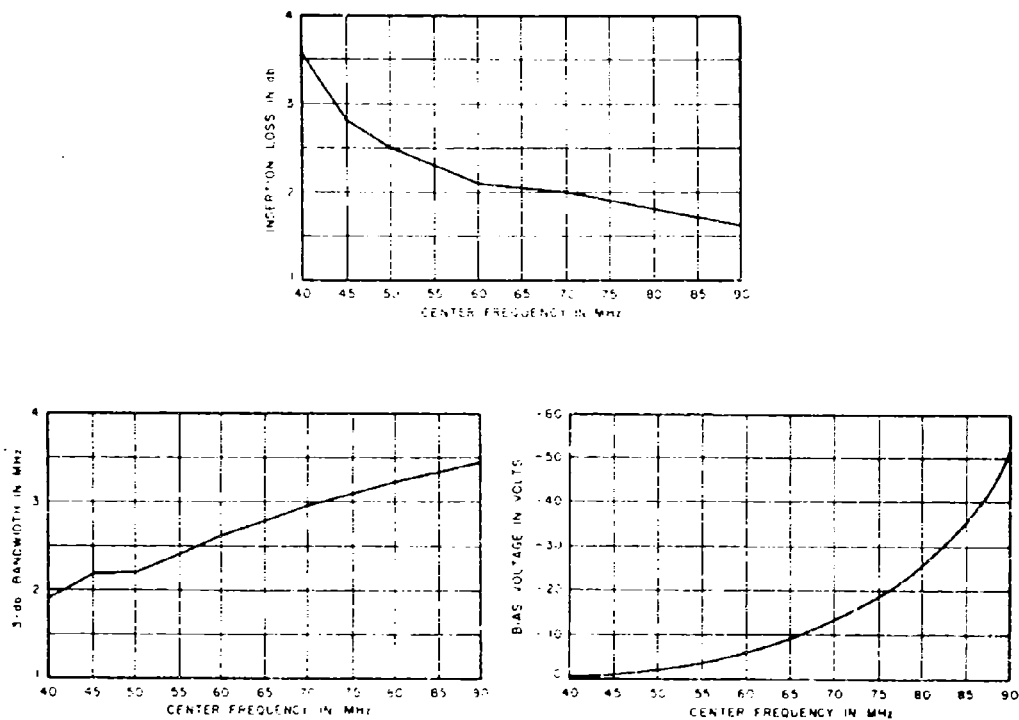
FIGURE 8. EXPERIMENTAL TWO-RESONATOR TUNABLE FILTER

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FIGURE 9. ELECTRICAL CHARACTERISTICS OF TWO-RESONATOR
TUNABLE FILTER

IV. RFI PROPERTIES OF VARACTOR-TUNED FILTERS

Several interference properties associated with varactor-tuned filters were investigated. Harmonic generation, particularly the second harmonic, is of importance for filters with tuning ranges of two to one or more. For such filters, the second harmonic of an input signal can appear within the tuning range, and if it is not suppressed properly, the harmonic becomes an erroneous signal. Perhaps of even greater importance is the third-order intermodulation product. This effect is very significant because it can occur at any frequency that the filter is tuned to. Other effects, such as cross modulation and detuning, can also occur and were investigated.

A. MEASUREMENT SETUP AND PROCEDURES

1. SECOND HARMONIC MEASUREMENT

The setup used to make harmonic measurements is shown in Figure 10. With the signal generator tuned to frequency f_0, the signal is passed through the low-pass filter to attenuate the second harmonic of the generator. Now with f_0 being the only significant signal frequency appearing at the filter input, the second harmonic of f_0 generated in the filter can be measured. The output of the filter is fed through a narrow band-pass filter so that the fundamental frequency does not appear at the input of the receiver. By calibrating the receiver, the level of the second harmonic can be obtained.

2. THIRD-ORDER INTERMODULATION MEASUREMENT

The setup used to make third-order intermodulation measurements is shown in Figure 11. With the filter tuned to f_0, the two signal generator frequencies (f_1 and f_2) are selected such that $2f_1 - f_2 = f_0$. The two signals (f_1 and f_2) are combined in the summing network and fed into the filter. Due to the nonlinear characteristics of the filter, a signal results at f_0 that is then fed into a very selective filter,
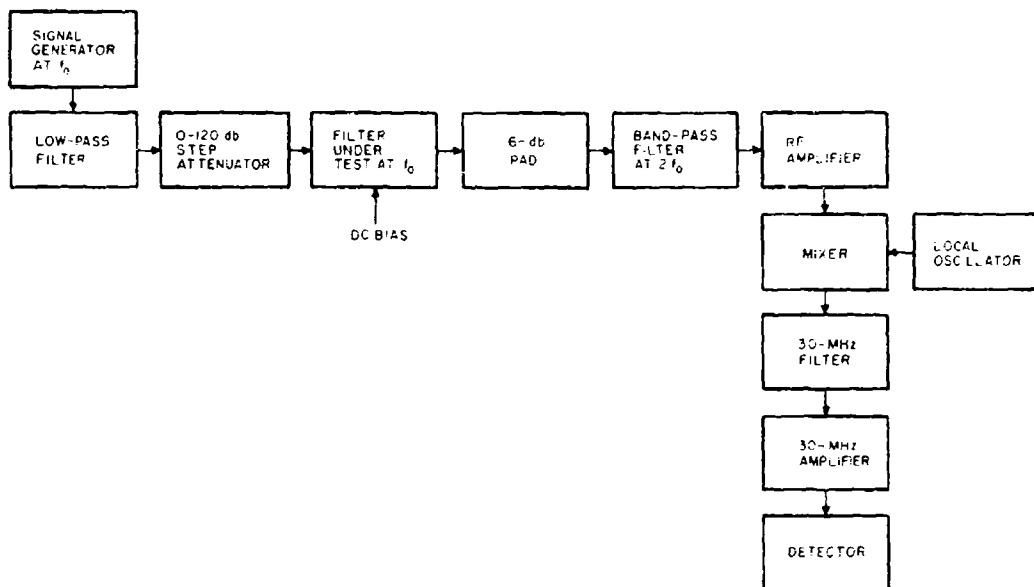
FIGURE 10.   TEST SETUP FOR SECOND HARMONIC
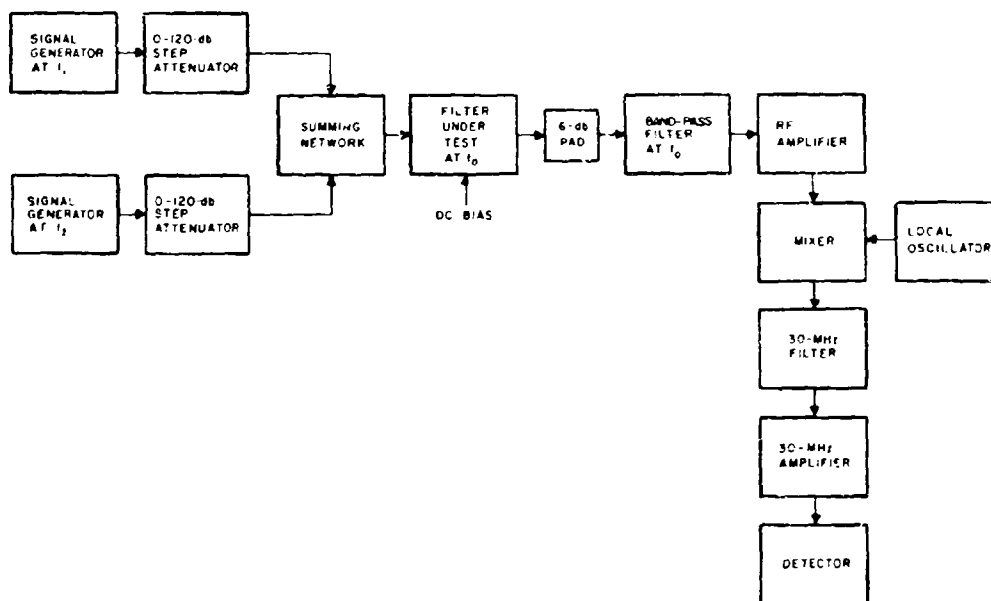   MEASUREMENTS ON TUNABLE FILTERS

FIGURE 11.   TEST SETUP FOR THIRD-ORDER INTERMODU-
             LATION MEASUREMENTS ON TUNABLE FILTER

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which rejects the frequencies f_1 and f_2. The level of the third-order intermodulation product at f_o is then measured with the receiver.

3.   CROSS-MODULATION MEASUREMENT

The test setup used to make cross modulation-measurements is shown in Figure 12. With signal generator no. 1 modulated to 100 percent by a 1-kHz signal and with signal generator no. 2 off, a reference corresponding to 100 percent cross modulation is obtained on the SWR indicator. The modulation is then removed from signal gene   tor no. 1 resulting in a CW input at f_1 into the filter. Signal generator no. 2 is now turned on and 100 percent modulated at 1 kHz. This signal is also fed into the filter under test. The output of the filter is then passed through a narrow band-pass filter tuned to f_1 and the 1-kHz modulation that is now present on the originally unmodulated signal is detected and read on the SWR indicator. The difference between the two readings obtained on the on the SWR indicator in db denotes how much modulation with respect to a 100 percent modulated signal is present on the desired CW signal and is called the
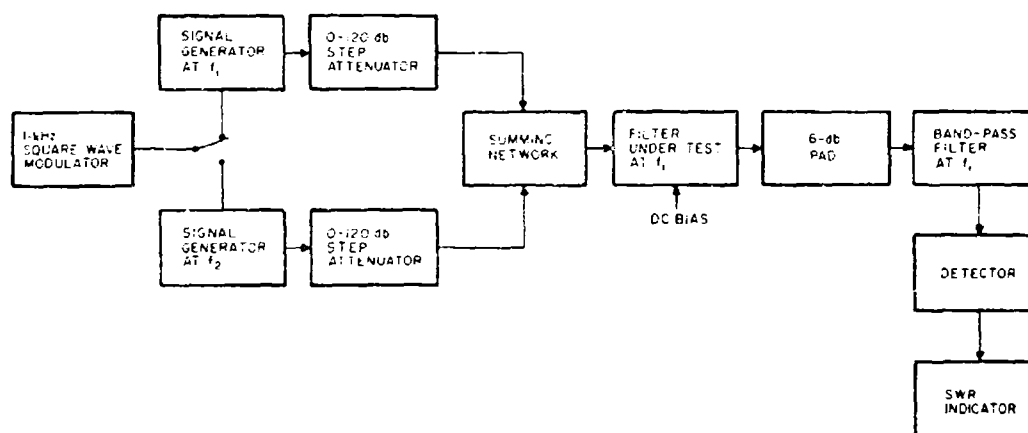

FIGURE 12.   TEST SETUP FOR CROSS-MODULATION MEAS-
UREMENTS ON TUNABLE FILTER

cross-modulation ratio. The cross-modulation ratio M is defined by the equation

$$M = 10 \log \frac{M_k}{M_I}$$

where

M_k = cross-modulation index--that is, the modulation index appearing on the desired signal

M_i = modulation index of the interfering signal

4.   DYNAMIC RANGE OR DETUNING MEASUREMENT

The test setup used to make this measurement is shown in Figure 13. The measurement consists of feeding a signal at the frequency that the filter is tuned to and recording the output level of the filter. As the input signal level is increased, the average capacitance value of the varactor decreases causing the center frequency of the filter to increase. This effect results in an increased transmission loss through the filter since the signal is now on the skirt of the filter selectivity characteristic. Thus, by plotting an input-output characteristic, a saturation or limiting effect can be observed.
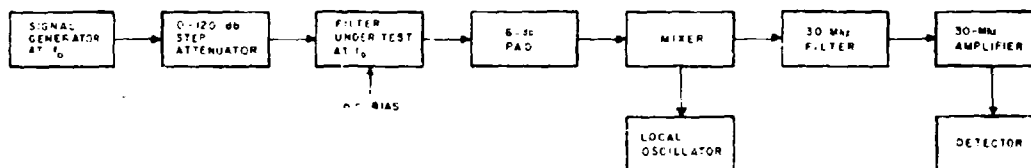
FIGURE 13.   TEST SETUP FOR DYNAMIC RANGE MEAS-
              UREMENTS ON TUNABLE FILTER

B.   RFI MEASUREMENTS OF SINGLE-RESONATOR FILTER

The measurements described in Section IV-A were made on the single-resonator filter described in Section II-B. The results are shown in Figures 14, 15, 16, 17, and 18.

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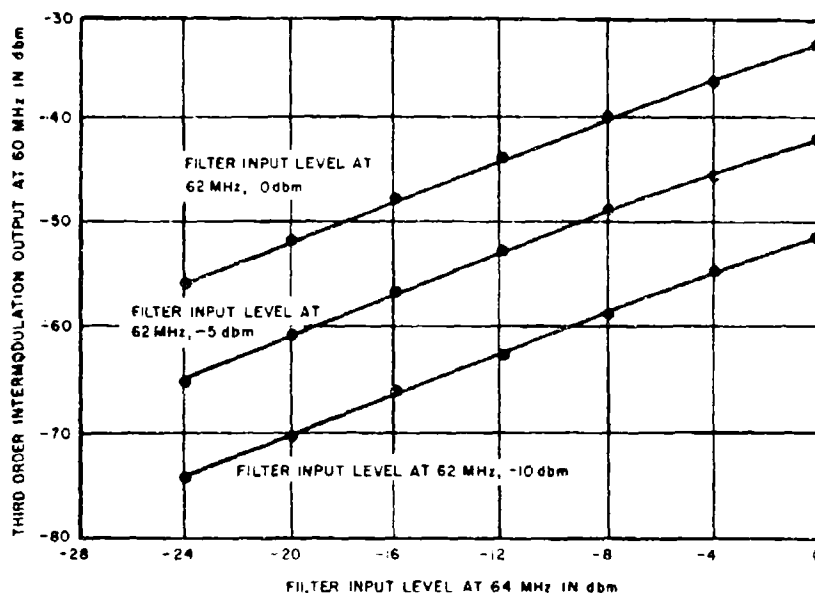
FIGURE 14. THIRD-ORDER INTERMODULATION OUTPUT VS
FILTER INPUT LEVELS FOR SINGLE-RESONATOR
VARACTOR-TUNED FILTER

1. THIRD-ORDER INTERMODULATION

It can be seen from Figure 14 that an intermodulation prod-
uct of -32.5 dbm can occur at 60 MHz when two interfering signals of
0 dbm at 62 and 64 MHz are present. This fact, however, does not
make the filter useless, since in most receiver applications such high-
level signals are rarely encountered.

From the expression relating third-order intermodulation
output level to the input signal levels, we can predict what the third-
order intermodulation output would be for input levels of -20 dbm at
62 and 64 MHz. The expression is

$$P_{12} = P_1 + 2P_2 + K_{12}$$

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where

P_{12} = third-order intermodulation output level, in dbm

P_1 = input level at f_1, in dbm

P_2 = input level at f_2, in dbm

K_{12} = constant depending on the device, in dbm
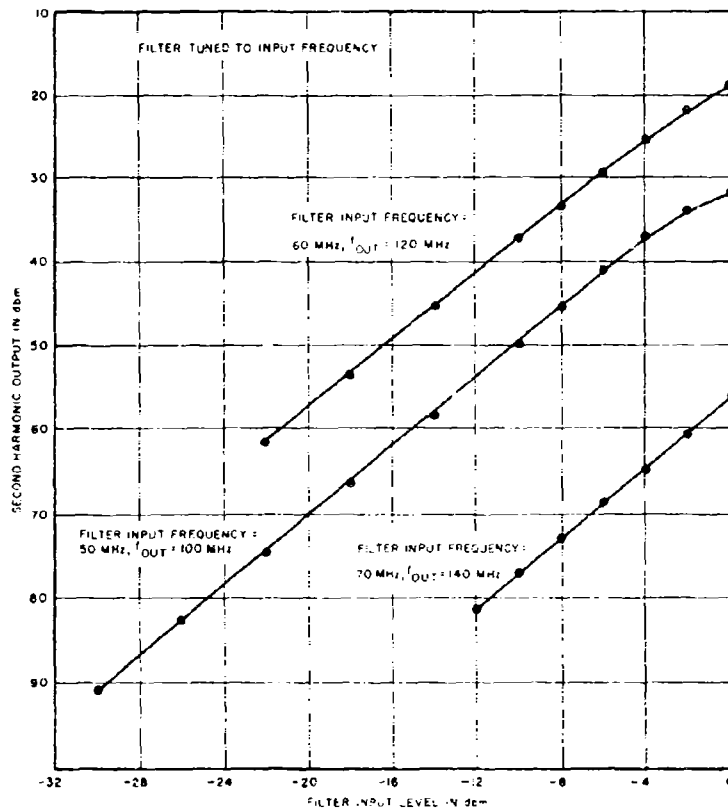
FIGURE 15. SECOND HARMONIC OUTPUT VS FUNDA-
MENTAL INPUT LEVEL FOR SINGLE-
RESONATOR VARACTOR-TUNED FILTER
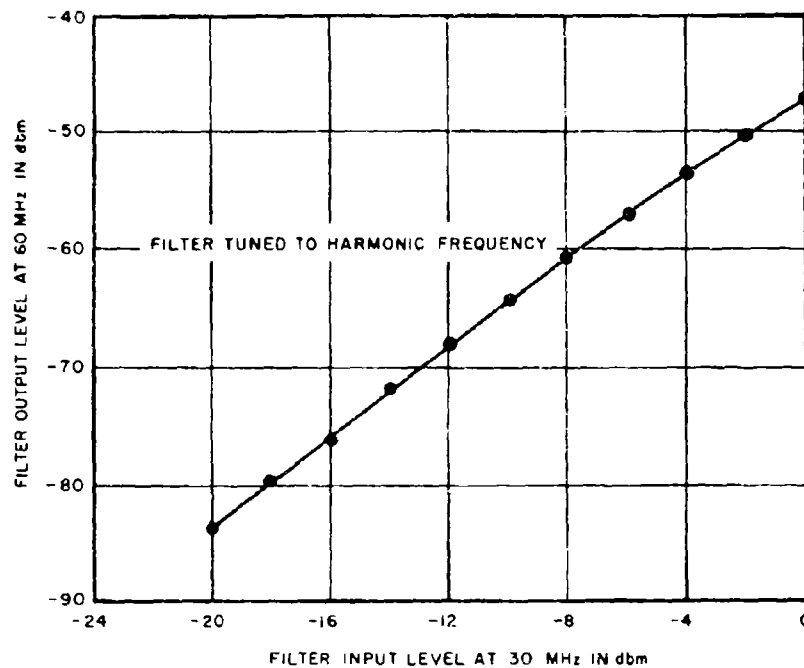
FIGURE 16. SECOND HARMONIC OUTPUT VS FUNDAMEN-
TAL INPUT LEVEL FOR SINGLE-RESONATOR
VARACTOR-TUNED FILTER

From our measurements, made with both input signals at 0 dbm, we have

$$-35.2 = 0 + 0 + K_{12}$$

$$K_{12} = -32.5 \text{ dbm}$$

Therefore

$$P_{12} = P_1 + 2P_2 - 32.5$$

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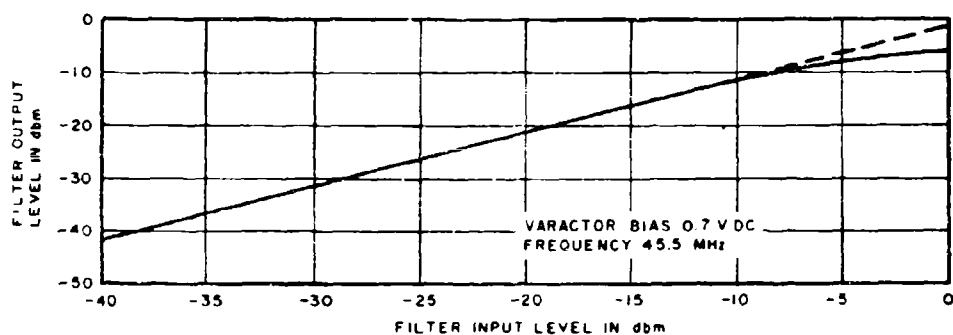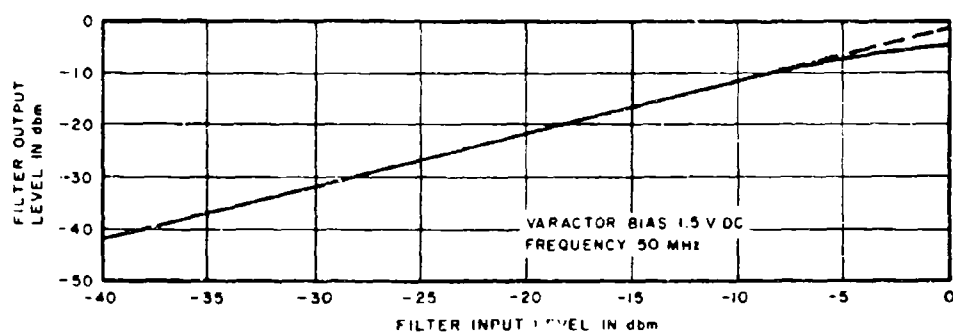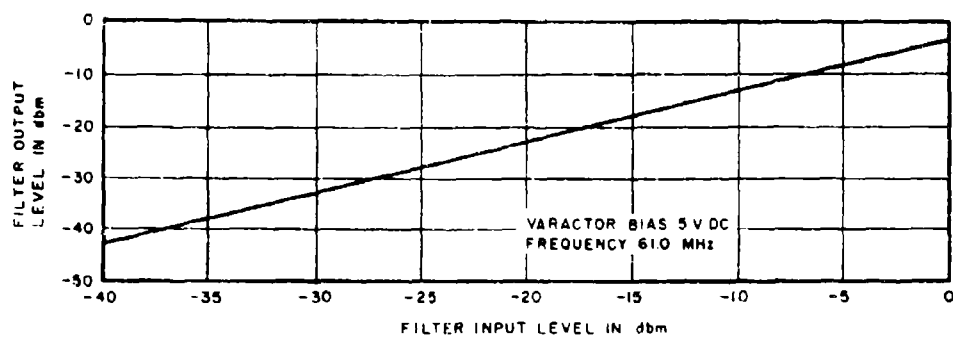
FIGURE 17. INPUT-OUTPUT CHARACTERISTICS AS A FUNC-
TION OF DC BIAS FOR SINGLE-RESONATOR
VARACTOR-TUNED FILTER

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For input levels of -20 dbm

$$P_{12} = -20 - 40 - 32.5 = -92.5 \text{ dbm}$$

Thus, we see that for two input signals at 62 and 64 MHz of -20 dbm each, the third-order intermodulation output level is -92.5 dbm. Under such conditions, operation of the filter as a preselector is possible. If lower third-order intermodulation levels are required, either narrower bandwidth or additional resonators should be used to attenuate the interfering signals and thereby reducing the inter-modulation level.
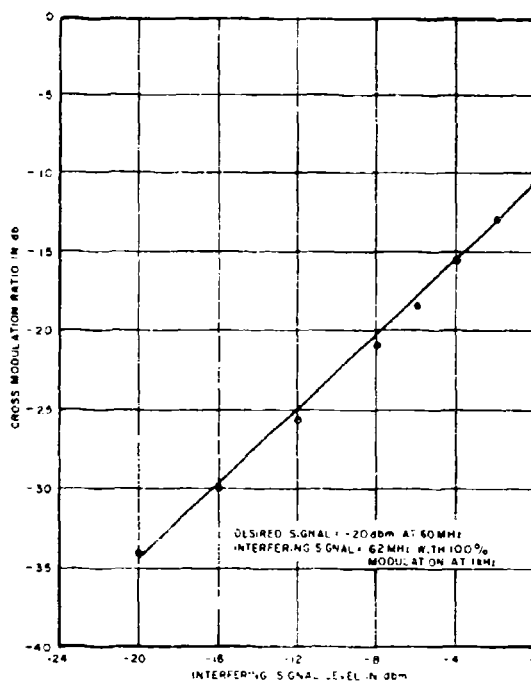
FIGURE 18.   CROSS-MODULATION RATIO VS INTERFERING
SIGNAL LEVEL FOR SINGLE-RESONATOR
VARACTOR-TUNED FILTER

2.   SECOND HARMONIC

Figure 15 shows the level of the generated second harmonic occurring at the output of the filter as a function of the fundamental frequency input power level.   The measurements were performed for input frequencies of 50, 60 and 70 MHz with the filter
tuned to these frequencies.   The output frequencies were 100, 120,
and 140 MHz, respectively.   It is seen that for input levels on the
order of 0 dbm at 60 MHz, a harmonic level of -20 dbm can occur
at 120 MHz.   In superheterodyne receiver applications, the seriousness of generating a second harmonic will depend upon the specific
frequency relationships between the RF, IF and LO frequencies.
For example, consider an up-converter where the IF is at 235 MHz
and the LO is at 175 MHz (for an RF at 60 MHz).   The second harmonic of the RF mixing with the second harmonic of the local oscillator produces an output at 230 MHz, which may well be in the IF
pass band.   If we have an IF of 10 MHz with an LO at 50 MHz, then
the 120 MHz (second harmonic) produces an output at 70 MHz with the
fundamental of the LO and at 20 MHz with the second harmonic of the
LO.   Thus, for the IF at 10 MHz, the spurious response is at twice
the IF and should not be a problem.

The filter property that is of more concern is the second
harmonic level occurring at the center frequency of the filter when a
signal, whose frequency is one half of the filter center frequency, is
applied at the input of the filter.   This measurement was made with
the filter centered at 60 MHz and the result is shown in Figure 16.
From this curve, it is seen that the generated output increases proportionally as a function of the input reaching a maximum of -47 dbm
for an input level of 0 dbm.

3.   DYNAMIC RANGE OR DETUNING

Figure 17 shows input-output characteristics of the filter
for DC bias voltages of 0.7, 1.5, and 5.0 volts applied to the varactor.
It can be seen that with 0.7 volt bias on the varactor, a signal whose

level is 0 dbm detunes the filter so that the signal appears on the skirt at approximately the lower 5-db point. With higher DC bias, the detuning decreases and it can be seen that with 5 volts bias, linear operation of the filter is obtained up to at least 0 dbm input signals. It is, therefore, advantageous to operate at higher bias-voltage levels when large signals are expected to be fed into the filter.

4. CROSS MODULATION

Figure 18 shows the cross-modulation ratio for the filter when a midband 60-MHz desired signal of -20 dbm and a 62-MHz interfering signal modulated to 100 percent with a 1-kHz signal are applied to the filter. With a 0-dbm interfering signal, the cross modulation is approximately 10 percent.

C. SINGLE-RESONATOR FILTER USING OPPOSINGLY BIASED VARACTORS CONNECTED IN PARALLEL

By using a pair of opposingly-biased matched-varactor diodes connected in parallel as shown in Figure 19A, a composite even functioned capacitance versus signal voltage characteristic as shown in Figure 19B is obtained. By performing the operations

$$Q = \int C(v)\, dv$$

and

$$i(t) = \frac{dQ}{dt}$$

It can be shown (reference 2) that the resulting Fourier Series expression for the current through the diode contains only odd harmonics of the input frequency. Thus, by using this diode configuration in the filter, the second harmonic response may be entirely eliminated.
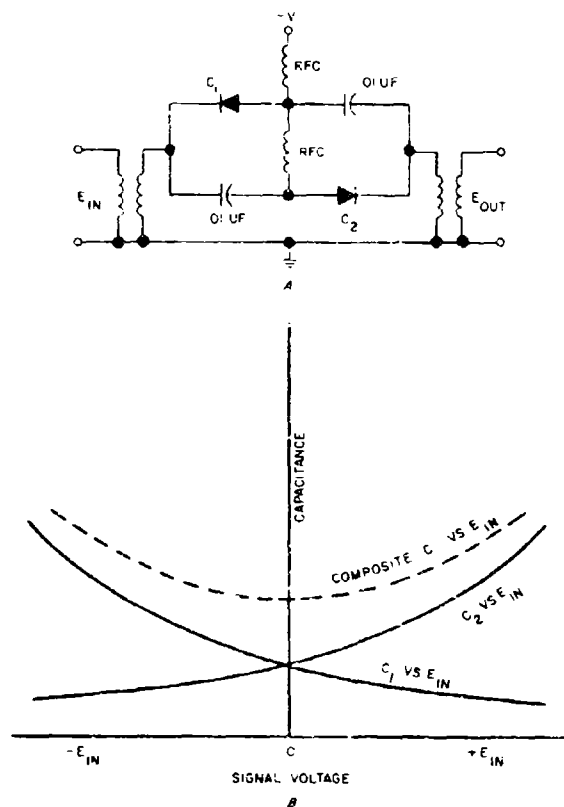
FIGURE 19. SINGLE-RESONATOR FILTER USING OPPO-
SINGLY BIASED VARACTORS CONNECTED
IN PARALLEL

This diode configuration was used in the single-resonator
filter for evaluation.   The configuration consisted of two similar diodes
with nearly the same characteristics as the diode previously used.
Since the total capacitance in the resonant circuit is now doubled, it
is expected that the minimum and maximum frequencies of the tuning
range will be reduced by a factor of 1.41.   The characteristics of the
filter are shown in Figure 20.
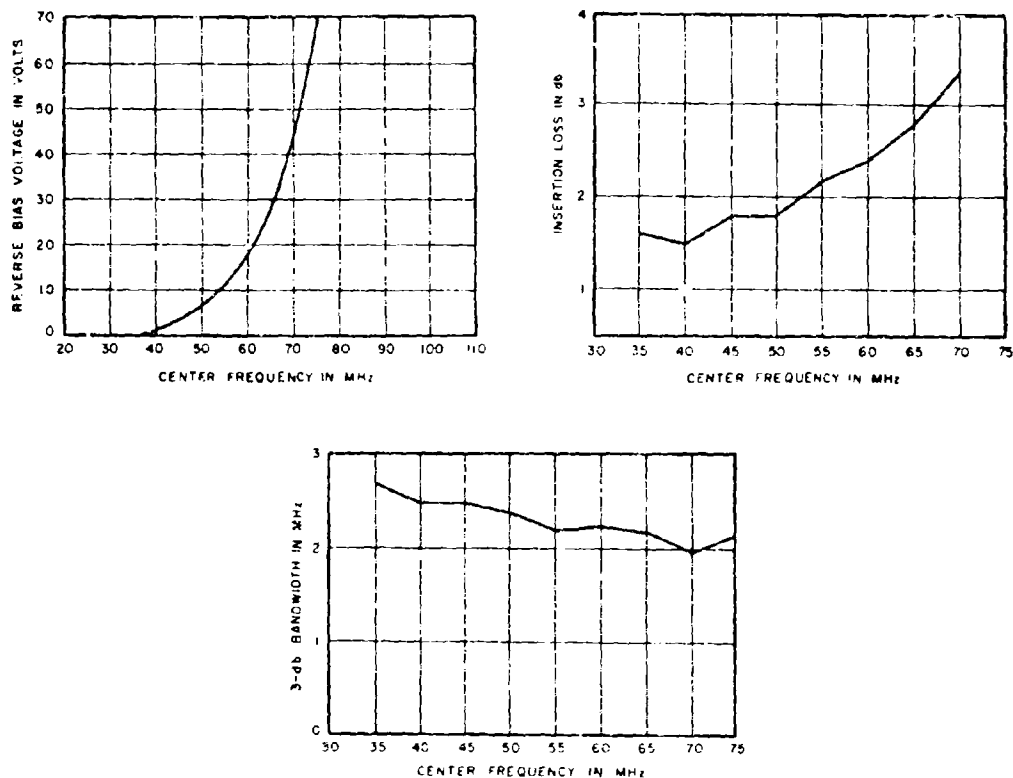
FIGURE 20. ELECTRICAL CHARACTERISTICS OF SINGLE-
RESONATOR FILTER USING OPPOSINGLY
BIASED VARACTORS CONNECTED IN PARALLEL

A direct comparison of the intermodulation curves of Fig-
ure 21 to the curves of Figure 14 cannot be made since the varactor
voltage is different for each filter and the third-order intermodulation
level is related to the operating point of the diode. However, if the
operating points were the same, a 3-db improvement in third-order
intermodulation (3-db lower) would result from the use of two diodes
since each diode handles only half of the power. The improvement
in third-order intermodulation that we obtained by the use of two

diodes is approximately 10 db, which tells us that the operating point of this filter when tuned to 60 MHz was such as to produce approximately 7-db lower third-order intermodulation power than the operating point of the single-diode filter when tuned to 60 MHz. In any case, there is a moderate improvement in third-order intermodulation when the opposingly-biased parallel varactors are used.
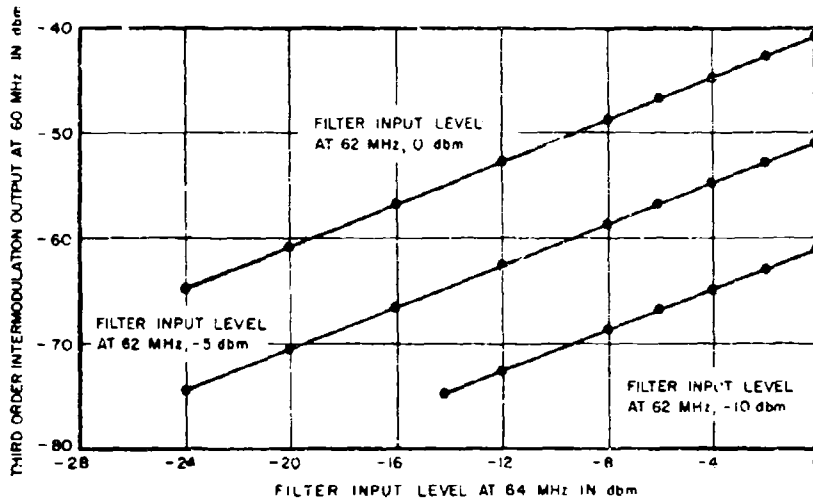
FIGURE 21. THIRD-ORDER INTERMODULATION OUTPUT VS FILTER INPUT LEVELS FOR SINGLE-RESONATOR VARACTOR-TUNED FILTER USING OPPOSINGLY BIASED VARACTORS CONNECTED IN PARALLEL

From Figure 22, we can see that for bias voltages equivalent to those used on the single diode filter, the "push-pull" filter has a 3-db greater dynamic range.

Second harmonic measurements were made on the filter with the test setup of Figure 10. The results are shown in Figure 23. It can clearly be seen that there is a significant improvement in second harmonic suppression when comparing these curves to those of
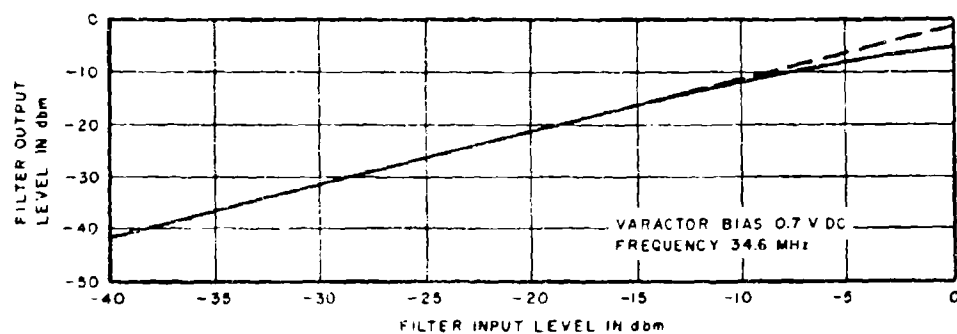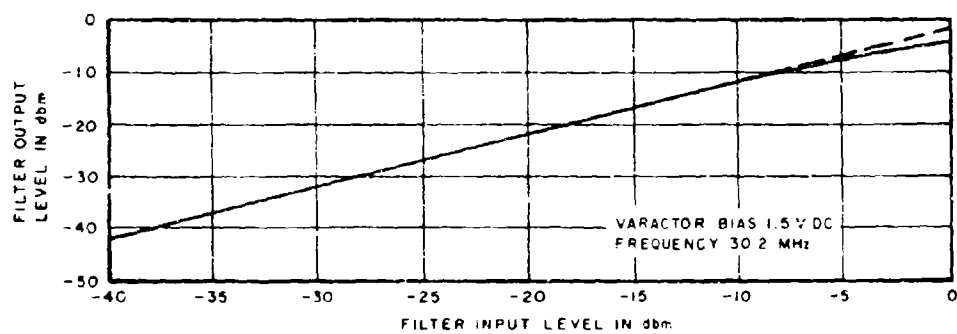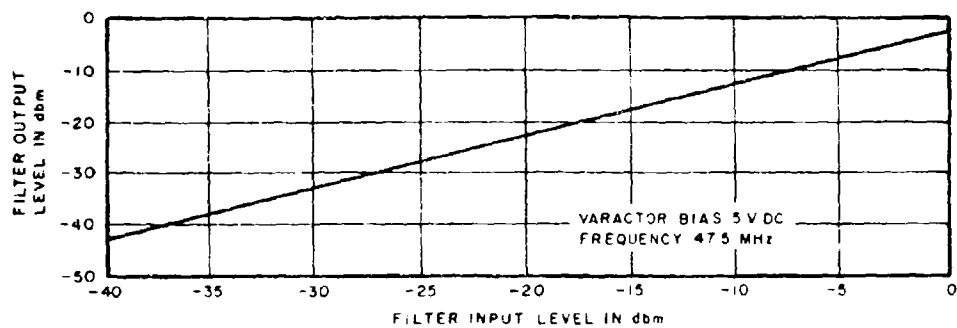
FIGURE 22. INPUT-OUTPUT CHARACTERISTICS AS A FUNC-
TION OF DC BIAS FOR SINGLE-RESONATOR VA-
RACTOR-TUNED FILTER USING OPPOSINGLY
BIASED VARACTORS CONNECTED IN PARALLEL

4-15
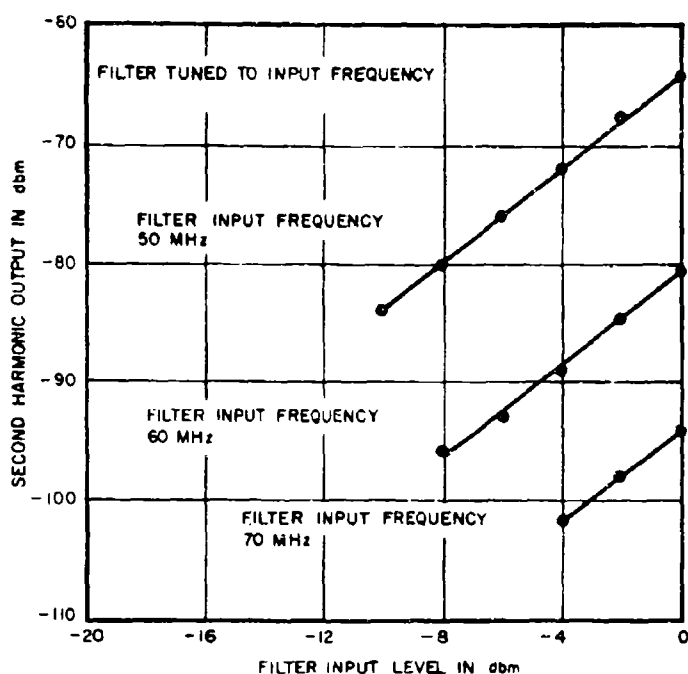
FIGURE 23. SECOND HARMONIC OUTPUT VS FUNDAMENTAL
LEVEL FOR SINGLE-RESONATOR VARACTOR-TUNED
FILTER USING OPPOSINGLY BIASED VARACTORS
CONNECTED IN PARALLEL

Figure 15. Typically, at 60-MHz input frequency, the second har-
monic level is -82 dbm for a 0-dbm input level. The other filter had
a second harmonic level of -18 dbm for 0-dbm input signal at 60 MHz.

With a signal at a frequency of one half the center frequen-
cy of the filter, the second harmonic could not be measured for input
levels as great as +7 dbm at 30 MHz. Since the sensitivity of the
measurement setup is -100 dbm, we can conclude that the second har-
monic of the input signal generated in the filter was at least 107 db
below the signal level.

. . .

D.   TWO-RESONATOR FILTER USING OPPOSINGLY BIASED
       VARACTORS CONNECTED IN PARALLEL

       The two-pole filter described in Section III-B was modi-
fied to incorporate the parallel-diode configuration.   The modified
filter characteristics are shown in Figure 24.
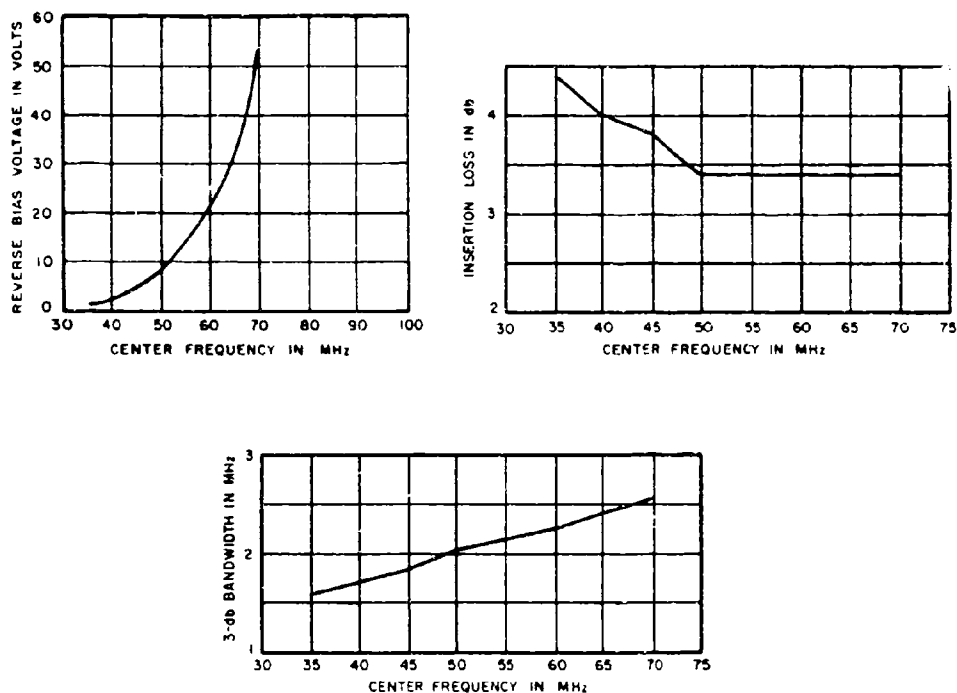
FIGURE 24.   ELECTRICAL CHARACTERISTICS OF TWO-
             RESONATOR FILTER USING OPPOSINGLY
             BIASED VARACTORS CONNECTED IN PARALLEL

       With the filter tuned to 60 MHz, the generated second
harmonic of the 30-MHz input signal was smaller than -107 dbm for
input levels as high as +7 dbm.   Thus, we see again that second har-
monic suppression by means of opposingly biased varactors in paral-
lel can be readily obtained.

Third-order intermodulation measurements were also made in this filter and the results are shown in Figure 25.
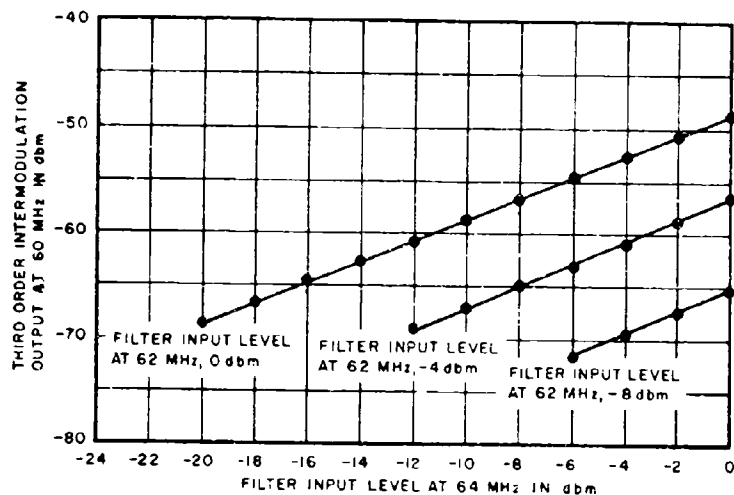
FIGURE 25. THIRD-ORDER INTERMODULATION OUTPUT VS FILTER INPUT LEVELS FOR TWO-RESONATOR VARACTOR-TUNED FILTER USING OPPOSINGLY BIASED VARACTORS CONNECTED IN PARALLEL

V. CONCLUSIONS

Voltage tunable filters using varactors as tuning elements have been found to have good RFI characteristics at VHF when used under the following restrictions:

1. The magnitude of the signal level should be smaller than the DC bias voltage to maintain a linear input-output characteristic at the signal frequency.

2. The tuning range should cover no more than an octave so that any generated second harmonic is out of the band of operation.

A typical application for a varactor-tuned filter which takes these restrictions into account is that of a receiver preselector.

When it becomes necessary to tune the filter over more than one octave, and it is required to minimize second harmonic generation, the configuration employing two opposingly biased varactors connected in parallel has been found particularly useful. Furthermore, this configuration also provides an additional 3-db reduction in third-order intermodulation for the same bias as compared to the single-diode configuration.

RFI measurements were taken on a single-resonator filter and a two-resonator filter, using a single varactor for each resonator. Comparison measurements were taken on the same filters, using opposingly biased varactors connected in parallel. A significant amount of additional second harmonic rejection was obtained. An extention of these results should apply to multiple-resonator filters.

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VI. REFERENCES

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2.   W. R. Avellino, "Efficient Odd Harmonic Generation with Opposingly Biased Varactor Diodes," Proc IEEE, p 868-869, July 1964.

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13. ABSTRACT

This report describes an investigation of circuit techniques to minimize spurious responses in varactor-tuned filters.

Design equations for the center frequency, bandwidth and insertion loss of single and multiple-resonator filters are presented. Series-resonant filters with one and two resonators were constructed and measurements of their performance characteristics were made.

One- and two resonator filters were also constructed using a pair of opposingly biased varactors (push-pull circuit) for each resonator. Measurements show that such a circuit can significantly reduce the second-harmonic generation and the response to a signal at one half the tuned frequency.

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